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COMSTAR Experiment:

The 19- and 28-GHz Receiving Electronics for the Crawford Hill COMSTAR Beacon Propagation Experiment

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This paper describes the receiving electronics built at the Bell Laboratories Crawford Hill facility at Holmdel, New Jersey to use the 19- and 28-GHz beacons on the COMSTAR satellites for propagation measurements. The receiving system accurately determines attenuation, differential phase, depolarization, bandwidth limitations and angular scatter of these signals produced by rain. This highly reliable system operates continuously and unattended; it automatically reacquires the beacon signals after dropout due to severe attenuation or momentary power outage. Correlations among strong and weak signal components are used to permit detection of weak cross-polarized signals during severe fading. Receiver noise bandwidths as low as 1.6 Hz are used. A high degree of phase stability is achieved in all circuits and components.

I. INTRODUCTION

The receiving electronics for the COMSTAR beacons has placed strong demands on technology in several areas in order to meet the requirements of the propagation experiments.[1,2,3] The receiving system built at the Bell Laboratories Crawford Hill facility includes a precision antenna[4] and the receiving electronics necessary to make maximum use of the 19- and 28-GHz beacon signals radiated by the COMSTAR satellites. The receiving electronics is the subject of this paper.

Continuous unattended operation is required so that all significant weather events are included in the resulting data base; thus, a very high

degree of reliability in the receiving electronics is required, and automatic reacquisition of the beacon signal after dropout due to severe attenuation or momentary power outage is essential. Since relative phases of the many signal components must be precisely measured, the phase stability of all circuits and components demanded careful attention. Also, circuit arrangements had to be devised to ensure that signals which were later to be compared in phase traversed a common path through high-gain amplifiers and other phase-sensitive equipment.

In order to obtain the maximum possible measuring range using the modest powers radiated by the satellite beacons, very narrow receiver noise bandwidths are required. Tbis puts a premium on the stability of the source oscillators in the satellites and the local oscillators in the earth stations. The receiver includes an AFC circuit with built-in memory to facilitate reacquisition after loss of signal. The feature also permits easy return to propagation measurements after use of the antenna system for radio astronomy studies during clear weather periods. Maximum use was made of known correlations among strong and weak signal components to permit detection of weak cross-polarized signals during severe fading. The following is an account of how these objectives were achieved, together with a description of the resulting apparatus and its operating characteristics. General design considerations are in Section II; Sections III and IV cover 19-GHz and 28-GHz receiver channels, respectively. Local oscillators and frequency control techniques are discussed in Section V. Section VI covers polarization switch synchronization. Data collection equipment is described in Section VII; Section VIII covers the receiver calibration source. Receiver performance and some sample data are included in Section IX.

II. RECEIVER DESIGN CONSIDERATIONS

The multichannel satellite beacon receiver must measure and record the signal amplitudes and phases listed in Table I for all rain events in order to satisfy the needs of the propagation experiment. Because of beacon[1] and rain[5] characteristics the receiver must: (*i*) have narrow (1.6 to 24 Hz) final noise bandwidths, (*ii*) keep the receiver frequencies within the narrow IF filter bandwidths (BW) as the beacon frequencies vary, (*iii*) hold local oscillator (LO) frequencies within a few Hz of their last known frequency for several minutes when rain attenuates the beacon signals below the frequency tracking threshold or when the primary power drops out, (*iv*) discriminate against the 1-kHz polarization switching sidebands while reacquiring the beacon signals automatically after long periods of loss of signal, (*v*) synchronize receiver polarization switches with the beacon polarization switch, and (*vi*) perform these functions reliably and automatically to permit continuous unattended operation. During clear air conditions the earth-station antenna[4] is used for radio astronomy

Table I — Signal amplitudes and phases

Receiver output	Description
Amplitude, 19 GHz	
A19H10	Horizontal copolarized,* 10-Hz IF BW
A19V10	Vertical copolarized,† 10-Hz IF BW
A19XH10	Horizontal crosspolarized,‡ 10-Hz IF aw
A19XH1	Horizontal crosspolarized,‡ 1-Hz IF BW
A19XV10	Vertical crosspolarized,†† 10-Hz IF aw
A19XV1	Vertical crosspolarized,†† 1-Hz IF BW
A19OA1	Off-axis receiving beam, 1-Hz IF aw
Phase difference, 19 GHz	
ϕ19V-H	(Vert. copol.†)-(horiz. copol.*), 10-Hz IF aw
ϕ19V-XV	(Vert. copol.†)-(vert. crosspol.††), 10-Hz IF aw
ϕ19V-XH	(Vert. copol.†)-(horiz. crosspol.‡), 10-Hz IF BW
Amplitude, 28 GHz	
A28V15C	Vertical copolarized,† 15-Hz IF aw, carrier
A28V1.5C	Vertical copolarized,† 1.5-Hz IF BW, carrier
A28V15U	Vertical copolarized,† 15-Hz IF BW, upper sideband
A28V15L	Vertical copolarized,† 15-Hz IF BW, lower sideband
A28XV15C	Vertical crosspolarized,†† 15-Hz IF aw, carrier
A28XV1.5C	Vertical crosspolarized,†† 1.5-Hz IF BW, carrier
A28OA1.5	Off-axis receiving beam, 1.5-Hz IF BW
Phase difference, 28 GHz	
ϕ28VC-XVC	(Vert. copol.† carrier)-(vert. crosspol.†† carrier), 15-Hz IF BW
ϕ28VC-U	(Vert. copol.† carrier)-(upper sideband), 15-Hz IF aw
ϕ28VC-L	(Vert. copol.† carrier)-(lower sideband), 15-Hz IF aw
Phase difference, crossband	
ϕ19V-28VC	(19-CHz vert. copol.,† 10-Hz IF BW) −(28-CHz vert. copol. carrier, 15-Hz IF BW)

* Horizontal copolarized is transmit horizontal and receive horizontal.
† Vertical copolarized is transmit vertical and receive vertical.
‡ Horizontal crosspolarized is transmit horizontal and receive vertical.
†† Vertical crosspolarized is transmit vertical and receive horizontal.

and other studies so the receiver must be easily restarted by people not intimately familiar with it.

Signal amplitudes and phases must remain accurate over receiver temperature changes of ±15°C and for signal attenuations of >30 dB. For example, amplitude and phase differences between the two copolarized 19-GHz channels must remain within 0.5 dB and 2°.[6] This requires careful attention to differential temperature control and to linearity.

Minimizing the number of frequency conversions is desirable to minimize receiver complexity and the number of spurious mixing products.

Since the 19- and 28-GHz beacon signals are derived from a common oscillator, they have the same frequency fluctuations. Thus, extended measuring range can be provided in the 28-GHz channels and in low signal 19-GHz channels (off axis and cross-polarization) if: (*i*) the corresponding 28-GHz and 19-GHz receiver LO sources are common, (*ii*) frequency fluctuations in the beacon and in LOs are tracked out with a common oscillator in a loop locked in frequency or phase to the 19-GHz vertically polarized signal, the signal that will experience the least at-

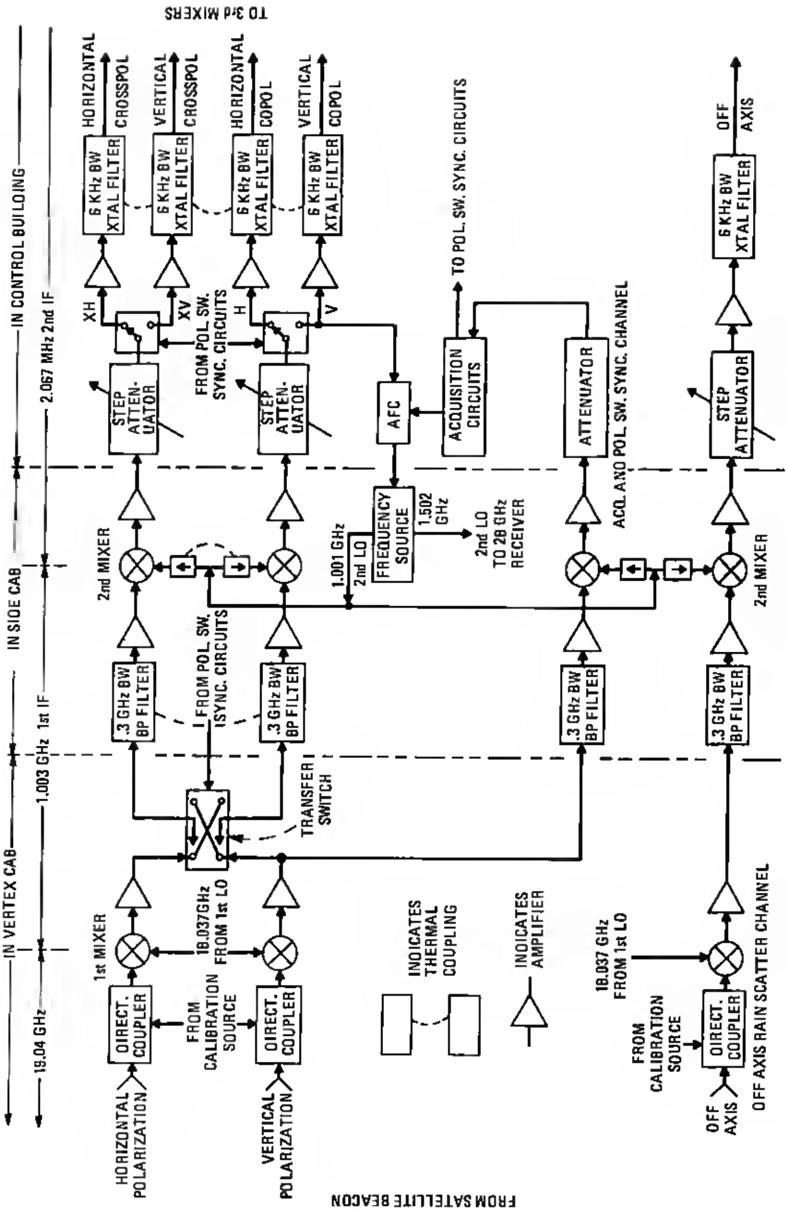


Fig. 1—19-GHz receiver channels: antenna feeds through second IF.

tenuation,[5] and (*iii*) very narrow band (~1 Hz BW) IF filters are used in the extended range channels. The provisions necessary for extending measuring range are easier to implement if all 19- and 28-GHz receiver IFs and LOs are in a 2:3 ratio.

The antenna has a small equipment room, the vertex cab, near the vertex of the main reflector, above both the azimuth and elevation axes; another room, the side cab, is to one side of the elevation axis but above the aximuth axis. The control building is about 15 meters (50 feet) away from the antenna at ground level (see Ref. 1, Fig. 3). The receiver is distributed among the three equipment locations to optimize noise performance and phase and amplitude stability, taking into account space limitations in the various equipment rooms.

The receiver is packaged by functional groups, e.g., bandpass filters, mixers and IF amplifiers for all the second frequency conversions are packaged together. This packaging approach maximizes differential phase and amplitude stability by minimizing temperature differential between similar components in different channels and also allows the receiver to be built in many separable blocks that may be "debugged" individually without complex interaction.

Since power line transients and momentary power outages are expected during heavy rain, all oscillators, filter stabilizing ovens and frequency memory registers are powered by batteries charged continuously from the power line.

III. 19-GHz RECEIVER CHANNELS

The 19-GHz receiver channels from the antenna feeds through the second IF crystal filters are shown in Fig. 1. The 19-GHz beacon signals are received with vertically and horizontally polarized feeds whose resulting antenna beams are pointed at the satellite. Another 19-GHz feed is located so its antenna beam is pointed toward an unoccupied synchronous orbit location about 0.74° off-axis from the satellite. This off-axis beam detects signals scattered by rain from the satellite beacon path into other potentially useful paths. This scattered signal is another possible source of cochannel crosstalk in multisatellite systems.[7,8]

Test signals from a calibration source described in Section VIII are fed into directional couplers between the antenna feeds and the first mixers. These calibrated test signals have adjustable amplitudes, are polarization switched and have adjustable simulated "crosspolarization" levels. The short term frequency stability of the calibration source is representative of the stabilities of the satellite beacons.

Signals are mixed with the 18.037-GHz first LO in the Schottky-diode balanced first mixers. Parametric amplifiers are not used because (*i*) they would contribute excessive amplitude and phase instability and (*ii*) adequate measuring range is more economically and easily provided by

narrow-band IF filtering. These mixers and associated low-noise broadband IF preamplifiers (~25 dB gain) are mounted on the feed horns. The first mixer and IF preamplifier single sideband noise figure (NF) of approximately 6.5 dB essentially determines the overall receiver NF. The remainder of the receiver contributes <0.1 dB. The first LO is distributed from a common source to all 19-GHz channels to preserve phase in the mixing process and to insure that the frequency fluctuations are correlated among the channels; isolating power dividers and ferrite isolators (see Fig. 5) insure >70 dB isolation between receiver channels through this common LO path. The 1.003 GHz first IF is dictated by the 1.056 GHz required bandwidth for the modulation sidebands in the 28 GHz receiver, by the need to keep 19- and 28-GHz IFs and LOs in a 2:3 ratio, and by the need to derive corresponding LOs from common sources.

After 30 dB more IF amplification the vertically and horizontally polarized channels are transfer-switched in synchronism with the beacon polarization switch. In this switch the copolarized signals (V and H) in adjacent time slots are switched into one receiver channel and the cross-polarized signals (XV and XH) are switched into the other channel. The accuracy of the differential amplitude and phase between the copolarized signals is very critical in the calculation of depolarization for other polarization angles.[6] The transfer switching insures that the phase and amplitude variations in most of the filters, amplifiers and long cable runs in the copolarized signal channel will affect the V and H signals identically. Thus, these variations will not affect the differential amplitude and phase measurements.

As indicated in Figure 1, the first mixers, first IF amplifiers and transfer switch are on the antenna feed assembly in the vertex cab. The mechanical rigidity of the feed assembly is adequate to ensure <0.5° differential phase variation (measurement limits) due to differential mechanical motion of feed components. Thermal differential phase variation for the feed assembly and associated receiver components is <0.2° per °C.

A separate 19-GHz receiver channel, used for signal acquisition and polarization switch synchronization, branches off before the transfer switch.

Cables run from the vertex cab to the side cab after the transfer switch but before 0.3-GHz BW bandpass filters (BPFs); these filters constrain the noise bandwidth to prevent noise from saturating the following broadband IF amplifiers and are mounted on an aluminum plate to minimize the temperature differential between them.

The second mixers reject image noise by more than 19 dB by phase-cancelling techniques (single sideband mixing) to permit the large fractional frequency spread between the first IF (1.003 GHz) and the

second IF (2.067 MHz). The 1.001 GHz second LO also is distributed from a common source to all 19-GHz channels. Isolating power dividers and ferrite isolators (see Fig. 6) insure >65 dB isolation between receiver channels through this common LO path. The second LO automatic frequency control (AFC) that follows average long-term frequency changes is described in Section V. The 5-MHz BW of IF amplifiers following the second mixers is narrow enough to prevent amplifier saturation on noise but wide enough both to preserve the rise time of the polarization switched signals and to render filter phase stability of little concern. The second IF amplifiers drive 50 meters of cable from the side cab to the control building.

Step attenuators match signal levels between the constant-gain low level (signals $S \leq$ noise N) front part of the receiver and the constant-gain high level ($S > N$) back part. This permits maximizing dynamic range by setting the clear air signal outputs near maximum. The second IF polarization switches separate the time-sequenced signals (V, H, XV, and XH) into separate channels. After further amplification, quartz crystal BPFs do the following: (*i*) further restrict the noise bandwidth, (*ii*) reject image frequency noise >20 dB for the third frequency conversion, (*iii*) significantly increase the rise time of the signal pulses and (*iv*) reduce second IF switch transients at the third conversion image frequency. The 6-kHz BW of these filters is as narrow as possible consistent with differential phase stability requirements. The four filters for the main beam channels (V, H, XV, and XH) are mounted together in a solid aluminum enclosure to minimize temperature differences.

The 2.067-MHz second IF frequency is a compromise restricted on the low side by LO noise close to the second LO frequency, by the need to preserve the rise time of the polarization-switched signals, and by the level of switching transients near tbe second IF frequency. The choice is restricted on the high side by phase stability considerations for the crystal BPFs and by the desire to minimize the number of receiver frequency conversions.

The acquisition and polarization switch synchronization channel preserves the 1-kHz polarization-switching signal from the beacon, uncontaminated by receiver switching. This channel is similar to the other receiver channels down to the crystal BPFs, except for the omission of switches. The acquisition and polarization switch synchronization circuits described in Sections V and VI contain their own automatic gain controlled IF amplifiers and crystal BPFs.

The off-axis rain scatter channel is essentially identical to the main beam channels except for the omission of polarization switches.

The 19-GHz receiver channels, continuing from the second IF crystal BPFs through the amplitude and phase detectors, are shown in Fig. 2. As indicated earlier, image rejection for the third frequency conversion

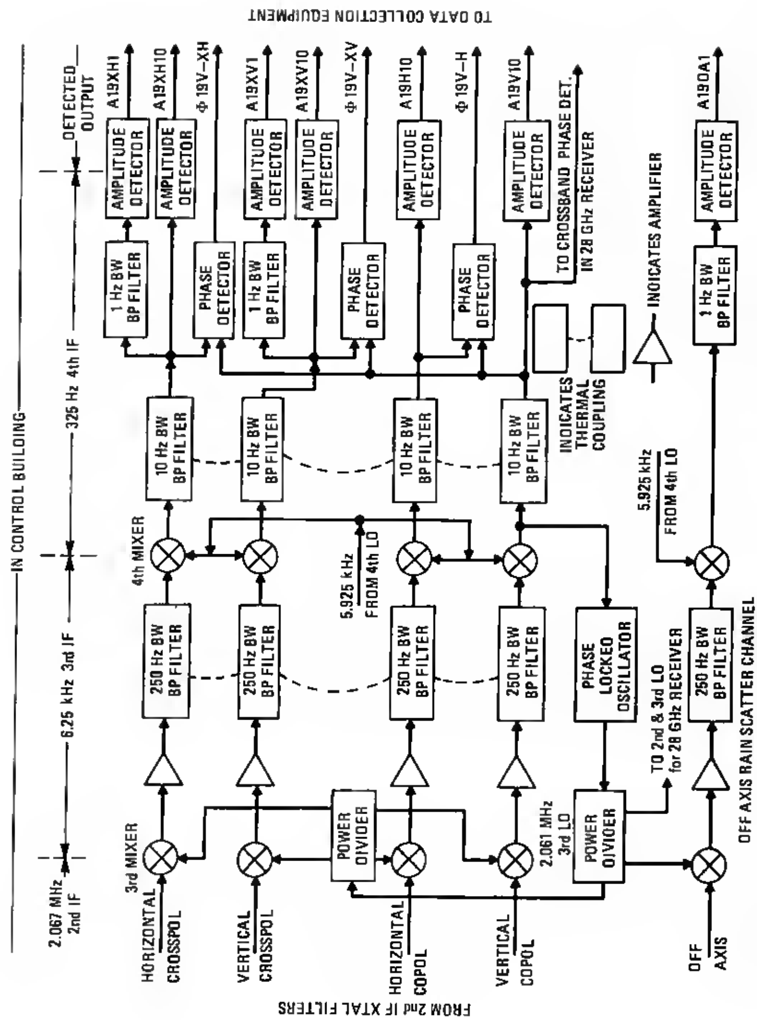


Fig. 2—19-GHz receiver channels: second IF through amplitude and phase detectors.

is done in the 6-kHz crystal BPFs. These stable high-Q filters permit a large fractional frequency spread between the second IF (2.067 MHz) and third IF (6.25 kHz). The 6.25 kHz third IF is constrained by phase stability considerations in the crystal BPFs and the need for at least 20 dB of filter rejection at the image frequency. The third mixers are double balanced (ring diode) with >30 dB signal to LO and LO to signal isolation. This isolation, along with the isolation in the LO power dividers (see Fig. 8), produces >80 dB isolation (measurement limit) between receiver channels through the common third LO path.

Because the third LO is phase-locked to the V copolarized signal channel using the phase locked loop (PLL) described in Section V, the short-term instabilities of the beacon oscillator and all receiver LOs are removed from all receiver channels at the third mixers. Thus, all third and fourth IF signals are as stable as the 325-Hz PLL reference as long as the PLL is locked.

Single resonator (single pole pair) active BPFs with 250 Hz BW follow 6.25 kHz IF amplifiers. These filters further restrict the noise BW, further reduce the polarization switching sidebands and receiver switching transients, and reject image noise and switching transients for the following fourth frequency conversion. The high input impedance of active analog multipliers used for the fourth mixers combined with the low output impedance of the operational amplifier supplying the 5.925-kHz fixed-frequency fourth LO (see Fig. 8) result in >80 dB isolation between receiver channels through the common fourth LO path.

Single resonator, 10-Hz 3-dB BW, active BPFs (~16 Hz equivalent noise BW) follow the fourth mixers. These BPFs are the final IF filters for the copolarized signal channels and for phase and amplitude measurement to moderate attenuation levels in the crosspolarized signal channels.

The specific third and fourth IFs are chosen to minimize the levels of the mixing products from the 1-kHz switching sidebands, both on the signal and the image sides of the third and fourth LOs. More than 60 dB filtering of the 1 kHz switching sidebands by the cascaded third and fourth IF BPFs insures that the phase ripple produced by the sidebands cannot exceed ±0.1°.

All of the third IF and 10-Hz fourth IF active BPFs and the fourth mixers are constructed with low-temperature-coefficient components. They also are enclosed in a temperature stabilized oven with <½°C internal temperature variation to maintain phase stability. The third and fourth IF filter Qs are nearly the same so that, assuming the same component variations, their contributions to the overall phase stability would be nearly equal. These filters were aged at oven temperature (65°C) before final alignment. The fourth IF BPF outputs are buffered through

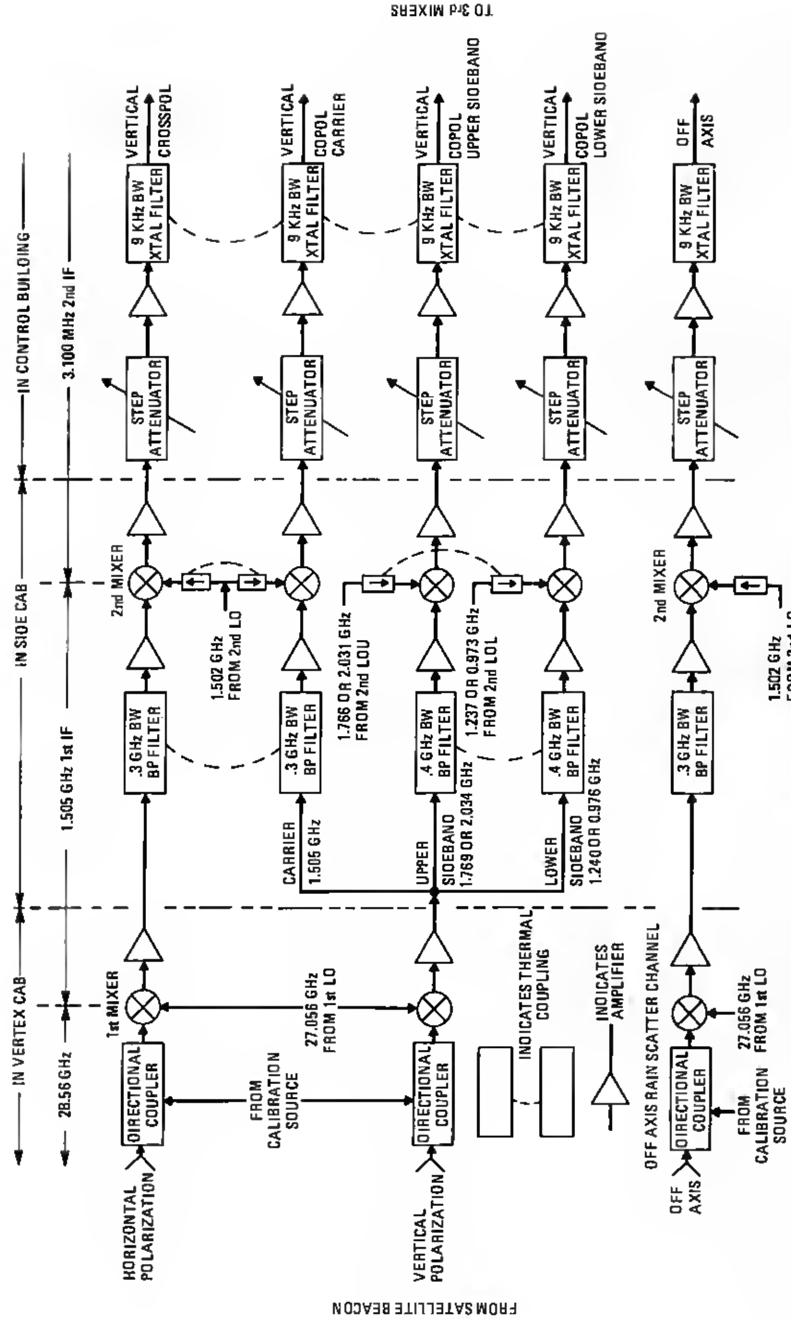


Fig. 3—28-GHz receiver channels: antenna feeds through second IF.

low-pass high-pass operational amplifiers that suppress low frequency ($1/f$) noise and broadband noise in the active BPF outputs.

Linear amplitude detectors following the fourth IF filters are full wave rectifiers comprising diodes in feedback paths of wideband low-drift operational amplifiers. These detectors are linear in dc output vs. signal input (voltage) to within ±0.1 dB over >60 dB signal range and a ±15°C temperature range. The phase detectors are commercial units with limiters (<±1° over 60 dB) in each channel followed by phase-to-pulse-width circuits and low-pass filter integrators. They measure phase unambiguously over 360°.

Because the PLL in tbe third LO removes short-term oscillator fluctuations from the fourth IF signal and because the XV, XH and rain scatter signal levels should always be less than the V copolarized signal level, the amplitude measuring range of the XV, XH, and rain scatter channels is extended by active BPFs with 1-Hz 3-dB BW. The bandwidth of these filters, which are similar to the 10-Hz BW filters and also are enclosed in an oven, is limited to about 1 Hz by the expected maximum rates of change of signal parameters.[2]

Again, the off-axis rain scatter channel is packaged separately and is essentially identical to the main beam crosspolarized signal channels except for the omission of the 10-Hz BW fourth IF BPF and the phase detector.

The outputs from the amplitude and phase detectors feed the dc signal-conditioning amplifiers in the data collection equipment in Fig. 16.

IV. 28-GHz RECEIVER CHANNELS

The 28-GHz receiver channels from the antenna feeds through the second IF crystal BPFs are shown in Fig. 3. The 28-GHz copolarized beacon signal (V) is received with a vertically polarized feed whose resulting antenna beam is coaxial with the 19-GHz beams. A horizontally polarized feed, whose beam is coaxial with the vertically polarized beam, receives the crosspolarized signal component (XV). These 28-GHz feeds share the main beam feed frame[4] with the 19-GHz main beam feeds. A 28-GHz off-axis rain scatter feed produces a beam coaxial with the 19-GHz off-axis beam.

Since many of the components and functions in the 28-GHz channels are similar to the corresponding ones in the 19-GHz channels, descriptions of similar components and functions are not repeated here. The major differences between the 19- and 28-GHz channels result because there is no polarization switching at 28 GHz but there are sidebands coherent with the 28-GHz carrier.[1] These sidebands (±264 MHz for two satellites and ±528 MHz for another) are used for measuring amplitude

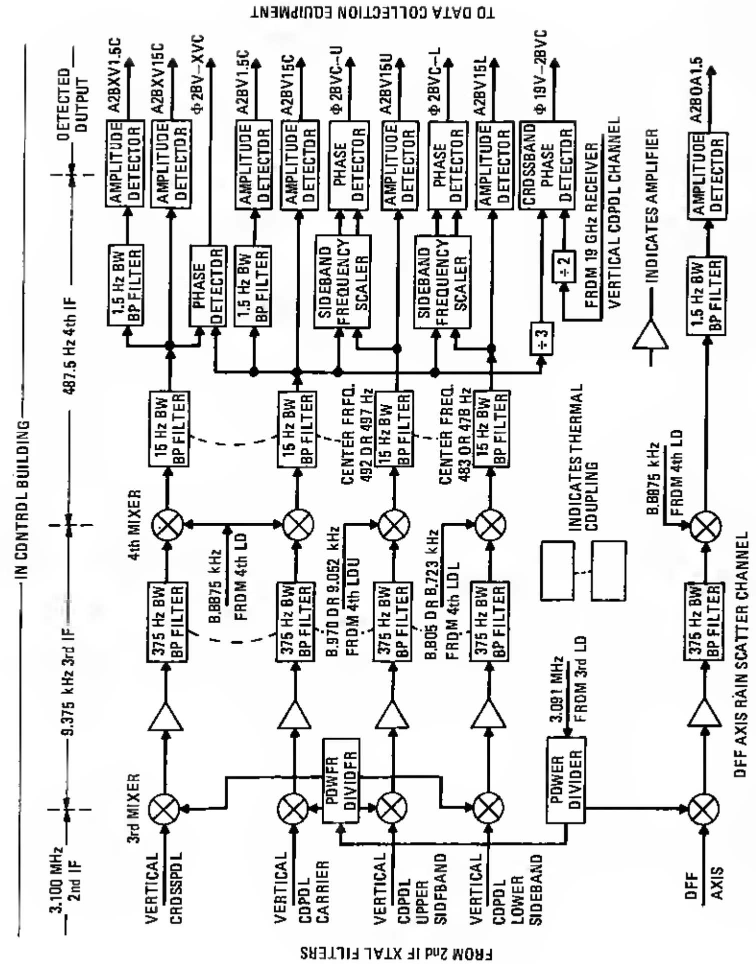


Fig. 4—28-GHz receiver channels: second IF through amplitude and phase detectors.

and delay dispersion.[1] Component locations discussed under 19-GHz receiver channels are indicated on Fig. 3.

As discussed earlier, 19- and 28-GHz receiver IFs and LOs are in a 2:3 ratio and corresponding LOs are derived from the same sources for extending measurement range and to permit measuring phase between 19- and 28-GHz carriers. This phase measurement yields the spectrum of east-west satellite pointing angle fluctuations because of the interferometer formed by the 2 meter (6.6 ft.) east-west separation between the 19- and 28-GHz satellite antennas. The measurement also yields dispersion information for a large frequency separation (9.5 GHz). In order to minimize effects of frequency changes on this phase measurement, all BPFs with significant delay (e.g., the crystal BPF) also have a 2:3 center frequency and bandwidth ratio between 19- and 28-GHz channels.

The calibration source provides calibrated test signals for all of the 28-GHz channels; sidebands coherent with the carrier and with either ±264 MHz or ±528 MHz separation can be selected.

A single fixed-frequency LO converts the entire 28.56 ± 0.528 GHz band, encompassing both sets of sidebands and the carrier, to the first IF centered at 1.5 GHz. This ensures that the short-term frequency stability of the most critical first LO is not compromised by other requirements. The first mixer and IF preamplifiers, which are similar to the 19-GHz units, have noise figures (SSB) of ≤7 dB over the entire 1.5 ± 0.528 GHz band. The first IF is a compromise that is pushed toward low frequencies by IF preamplifier noise-figure considerations and toward higb frequencies both by the need to pass the wide bandwidth and by first LO noise considerations. Since many standard octave bandwidth components are available for the 1- to 2-GHz band, the 28-GHz receiver first IF band is set at 1.5 ± 0.528 GHz. This constraint, along with the 2:3 ratio IF and LO constraint, fixes the first IF of the 19-GHz receiver at 1 GHz. The specific IF frequency, 1.504 GHz, results from maximizing the frequency separation between the desired IF signals and spurious mixing products from this and later frequency conversions. Isolation between receiver channels through the common first LO path (see Fig. 5) is >65 dB. Sidebands (SBs) are split into separate channels in the first IF so the bandwidth of the remainder of the receiver can be narrowed to increase sensitivity. (With the existing radio link parameters the maximum S/N in a 1-GHz BW is <-20 dB.) The upper sideband (USB) channel uses the same 0.4 GHz BW BPF, broadband first IF amplifiers, second mixer, second IF amplifiers and crystal BPFs for both the +264 and +528 MHz USBs. The corresponding parts of the LSB channel also are the same for both sideband frequencies. The same narrowband crystal filters are usable because the third LO SB frequency corrections, discussed in the next paragraph and in Section V, are done in the second

LOs. This second LO correction keeps the frequency spread between the different SBs small (<±200 Hz) at the second IF.

The measurement of delay dispersion[1] requires measuring phase between the carrier and the USB scaled by 108/109 or 108/110 and between the carrier and the LSB scaled by 108/107 or 108/106. These scaling factors are the exact ratios between the carrier and the SBs for the different satellites. Preserving the carrier and SBs in these ratios throughout the receiver requires separate LOs for carrier and SBs scaled similarly in frequency. Simple mixing by a single LO, as is done in the first frequency conversion, does not preserve the ratios. Therefore, sideband LO frequency corrections must be made elsewhere for this conversion and any other such conversion. The SB frequency corrections for the first, second, and third LOs are made to the SB second LO frequencies. This is done by adding or subtracting the "missing" portions of these LOs [(1 or 2)/108 $\times f_{LO}$] to the second LO used for carrier mixing. The LO implementations in Fig. 6 for making the corrections are described later. Provisions for remotely switching the SB channel LOs in the side cab from ±264 to ±528 MHz for the different satellites are provided in the control room. Isolation among the V carrier, XV and off-axis channel through their common LO path is >61 dB.

The 28-GHz off-axis rain scatter channel is identical to the 28-GHz XV channel and packaged together with the 19-GHz off-axis channel.

The 28-GHz receiver channels, continuing from the crystal BPFs through the amplitude and phase detectors, are illustrated in Fig. 4. Since the SB LO frequency correction for the third LO is included in the second LO, the third LO is common to all 28-GHz channels. Isolation among 28-GHz channels through the common third LO path (see Fig. 8) is provided in the same way as in the 19-GHz channels and is >80 dB (measurement limit).

In the USB channel the same broadband third mixer and third IF amplifier are used for both USB frequencies. The same is true for the LSB channel.

Although not shown separately in Fig. 4, the active third IF and fourth IF BPFs and fourth mixers are separate for the USB channels for the 264 and 528 MHz SBs. These components are separate also for the corresponding LSB channels. The fourth LOs for the four SB channels include the fourth LO frequency corrections (see Fig. 10). Separate channel filters are used at this point in the receiver because the different SB frequencies are spread approximately 1 percent in frequency (1/107 and 1/109) and the BPF 3-dB BWs are only 3 percent.

The 1.5-Hz BW BPFs in the XV and V carrier channels provide measuring range extension as described for 19-GHz receiver channels.

The active filters and fourth mixers for the XV and V carrier channels and for the ±264 MHz USB and LSB channels are in one oven. The ±528

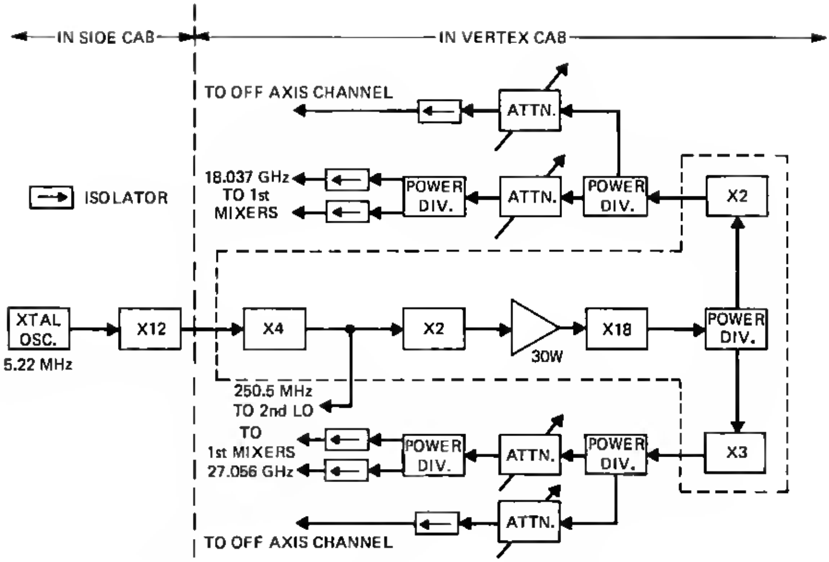


Fig. 5—Receiver first local oscillator (crystal oscillator frequency is 5.218991 MHz).

MHz USB and LSB channels and the 1.5-Hz BW active BPFs are in another oven.

In the SB frequency scalers, the SB frequencies are scaled by 107/108 and 109/108 or by 106/108 and 110/108 for the delay dispersion phase measurement. The sideband signals are limited and scaled by PLL frequency multipliers and digital dividers.

The crossband phase measurement is made between the 28.56- and 19.04-GHz V-copolarized carriers. The 325-Hz fourth IF signal from the 19-GHz receiver channel (Fig. 2) is divided by 2 and the 487.5 Hz fourth IF signal from the 28-GHz receiver channel (Fig. 4) is divided by 3 before the crossband phase detector (Fig. 4).

The off-axis rain scatter channel is similar to the XV channel. Both the 19- and 28-GHz off-axis channels are packaged together and are separate from the mainbeam channels.

V. LOCAL OSCILLATORS AND FREQUENCY CONTROL TECHNIQUES

Frequency requirements for the LOs and the receiver channel-to-channel isolation through the common LO paths are discussed in the previous sections.

5.1 First LO

The first LO is shown in Fig. 5. A 5.218991-MHz high stability (1×10^{-10}/day, 1×10^{-11}/0.1 sec) quartz crystal oscillator is multiplied first

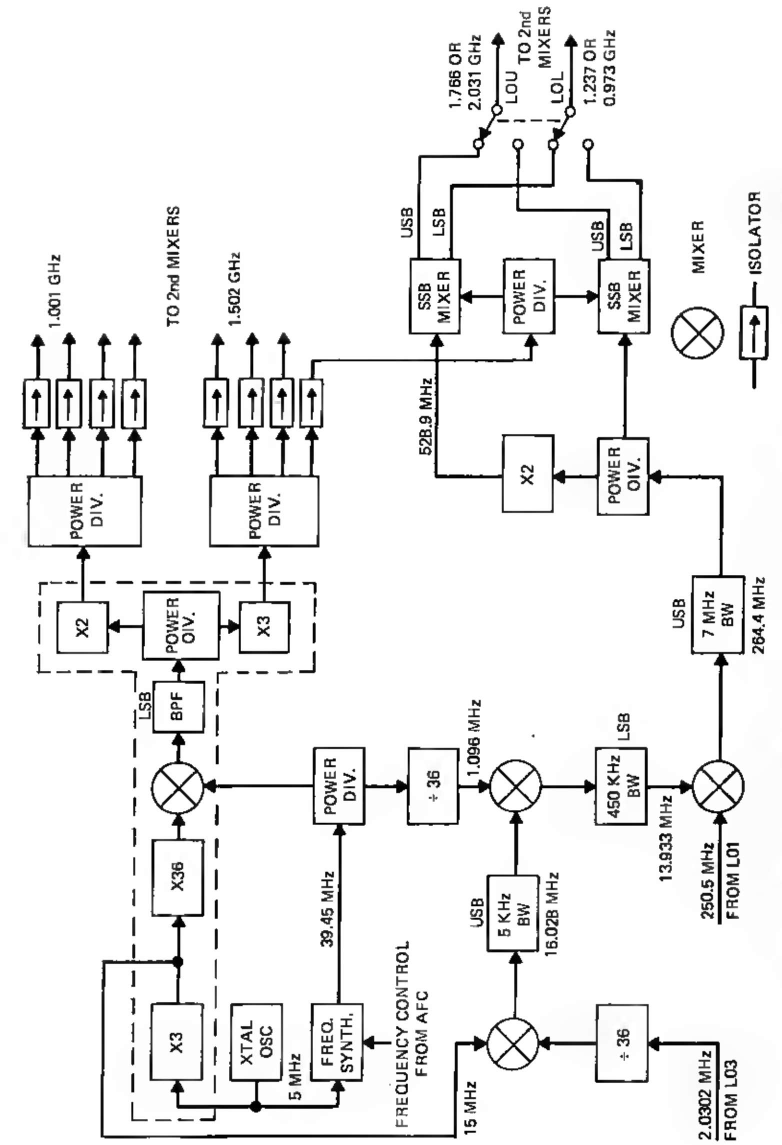


Fig. 6—Receiver second local oscillator in side cab (crystal oscillator frequency is 5.000000 MHz).

by 48 to 250.5 MHz for the second LO sideband correction frequency described earlier and then by 36 to 9.019 GHz; further multiplication by 2 or by 3 produces the coherent LO signals for the 19- and 28-GHz receiver channels. Passive varactor frequency multipliers follow the power amplifier at 501 MHz. The first LO does not include frequency control so the stability of the basic crystal oscillator is preserved. The first LO frequency multipliers from 62.6 MHz to 18 and 27 GHz, power dividers, isolators, and waveguide distribution are mounted on the antenna feed assembly in the vertex cab.

5.2 Second LO

The second LO is derived from a 5.000000-MHz quartz frequency standard as shown in Fig. 6. Multiplication by 3 produces a component of the sideband correction frequency at 15 MHz. This signal is multiplied by 36 to 540 MHz where the frequency of a 39-MHz frequency synthesizer is subtracted in a mixer. The synthesizer frequency is controlled by the automatic frequency control loop (AFC) indicated on Fig. 1 and described later in this section. Multiplication by 2 and by 3 after the mixer produces the coherent second LO signals for the 19- and 28-GHz receiver channels.

The second LO does the first, second, and third LO frequency scaling described earlier for the sideband channels of the 28-GHz receiver. This scaling starts by adding 15 MHz from the second LO frequency multiplier and 28.6 KHz from the third LO (i.e., 1.03 MHz ÷ 36) in a mixer; these are both 1/108 of the contribution of their respective sources to the second and third LOs. The synthesizer output divided by 36 to 1.096 MHz is then subtracted from the correction term since the synthesizer frequency itself is subtracted in the 1.562-GHz LO multiplier chain; this 1.096 MHz is also 1/108 of the synthesizer contribution to the 1.502-GHz LO. The resulting 13.933 MHz correction term, containing the contributions from the third LO and both components (oscillator and synthesizer) of the second LO, is then added to the 250.5 MHz = 27,056/108 MHz correction term from the first LO. The resulting 264.4 MHz correction term, $(1/108)(f_{LO1} + f_{LO2} + f_{LO3})$, is then added to the 1.502-GHz second LO frequency to produce the 1.766-GHz upper-sideband second LOU, described in the section on 28-GHz receiver channels. The correction term also is subtracted from the second LO frequency to produce the corresponding second LOL at 1.237 GHz.

The 264.4-MHz correction term is multiplied by 2 and similarly added and subtracted to provide second LOU and LOL for the beacons with ±528.8-MHz sideband separation.

The order of mixing of the correction frequencies makes the filtering easiest.

5.3 AFC loop

An AFC loop controlling the second LO is used to remove long-term frequency variations in both the satellite and the other receiver LOs. This AFC loop must meet the following requirements:

(*i*) Track over ±200 KHz.
(*ii*) Track a ±1.1 Hz/sec frequency ramp.
(*iii*) Maintain ±10-Hz accuracy while tracking.
(*iv*) Have a frequency averaging time of around 1 second.
(*v*) Hold frequency on command for 10 minutes to within ±50 Hz.
(*vi*) Contribute negligible short-term frequency instability.

Requirements (*i*) and (*ii*) are set by the expected long-term frequency behavior of the beacon.[1] Requirement (*iii*) guarantees that the phase-locked loop used for short-term frequency correction need not have a large capture range. Requirement (*iv*) constrains the AFC loop to average over most short-term frequency variations and track only long-term effects. This will improve the loop's low-SNR tracking performance and allow a better estimate of the signal's average frequency during outages. Requirement (*v*) states that the receiver frequency stability during signal outages of up to 10 minutes will be much better than the satellite's; thus, the frequency region which must be searched during reacquisition is determined only by the satellite.

These requirements impose severe stability and tunability requirements on the loop frequency generation element. A digitally controlled frequency synthesizer provides frequency memory and long-term stability equal to that of its reference oscillator and thus was chosen for this application.

It is a simple "fact of life" that a fixed-frequency oscillator has better short-term frequency stability than a variable-frequency oscillator, be it a VCXO or synthesizer. To prevent the synthesizer from adding to receiver short-term instability, a relatively low-frequency synthesizer is mixed with a high-frequency fixed source, as discussed previously. In addition, a synthesizer exhibiting low phase noise was chosen for this application. The exact synthesizer and first IF frequencies were chosen to insure that any internally-generated signal-frequency or image-frequency spurious signals would be well outside IF filter passbands.

The circuitry for controlling the synthesizer output frequency is shown in Fig. 7. The 2.067-MHz IF from the 19-GHz vertically polarized feed is filtered to remove the 12.5-kHz image response created by the following conversion. The filter output is levelled by an AGC amplifier to keep the discriminator input level constant. Since the clear-air S/N at the filter output is ~30 dB, the AGC will be noise-dominated only for

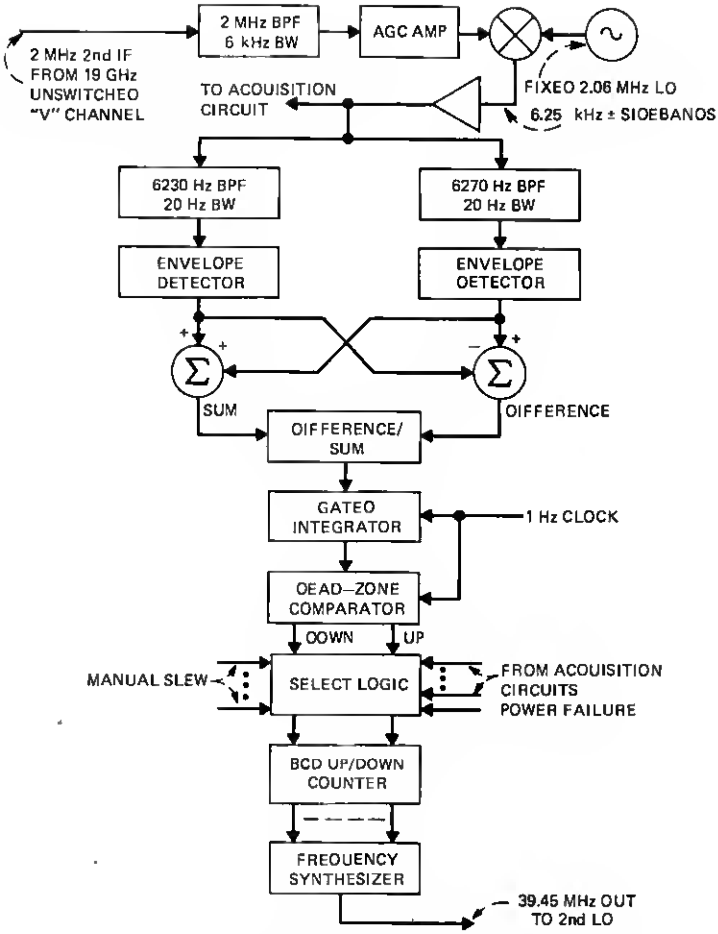


Fig. 7—Receiver second local oscillator AFC circuitry (digital portion).

fades of 30 dB or more, and will hold the signal component of its output constant for fades of lesser depth.

To obtain the desired narrow discriminator bandwidth, the signal is downconverted with a fixed-frequency LO to 6.25 kHz and fed to two 20-Hz-wide active bandpass filters tuned 20 Hz above and below center frequency. These filters are enclosed in a temperature-controlled oven to enhance their long-term frequency stability. The filter outputs are envelope-detected and the detector outputs summed and differenced. The difference is divided by the sum voltage in an analog divider. This division operation provides the effect of additional AGC and widens the capture range of the complete AFC loop by "propping up" the tails of the discriminator S curve. The 6.25-kHz signal, with 1-kHz polarization-

switching sidebands, is fed also to the acquisition detector described later.

The discriminator output is filtered by an integrator which is reset once per second. The dead-zone comparator produces an output pulse on one or the other of two lines if the integrator output indicates an average frequency error of more than 2 Hz. These pulses are used to increment or decrement a six-decade digital up-down counter; the BCD counter controls the synthesizer output frequency to 1-Hz resolution. Since the synthesizer frequency is doubled before use as a 19-GHz LO, an average signal-frequency error of greater than 2 Hz over a 1-second interval will produce a 2-Hz local oscillator frequency correction during the next second. The loop will tbus track a signal moving at up to 2 Hz/second to ±2-Hz accuracy. Witb six-digit frequency resolution, the loop tracking range is 2 MHz.

The up-down counter may also be controlled from other sources. During the signal acquisition phase the counter is automatically stepped up or down to sweep the receiver around the expected signal frequency. The counter may also be stepped up or down manually. These operations are discussed later in this section.

Long-term frequency memory is implemented by disabling the counter inputs when the signal fades below tbe threshold of the tracking loop. Power-failure protection is provided by powering the counter from an uninterruptible battery power source. In both cases the long-term "frequency memory" of tbe loop is solely determined by the frequency stability of the synthesizer master oscillator.

5.4 Third LO and PLL

The third LO is shown in Fig. 8. The 2.06-MHz voltage-controlled crystal oscillator (VCXO) source is controlled by the phase locked loop indicated on Fig. 2. The oscillator frequency is used directly for the 19-GHz receiver channels and after division by 2 and multiplication by 3 for the 28-GHz channels. The phase locked loop removes the remaining short-term oscillator frequency variations and presents a frequency-stable signal to the final 10-Hz and 1-Hz bandwidth receiver filters. This PLL must meet the following requirements:

- (*i*) Track over ±50 Hz.
- (*ii*) Have a loop bandwidth selectable around 10 Hz.
- (*iii*) Hold at "center" frequency on command to within ±2 Hz.
- (*iv*) Exhibit no frequency "walk-off," and recover quickly from signal outages.

Requirement (*i*) allows the loop to track the expected range of short-term frequency variations. Requirement (*ii*) permits the trading of loop threshold for amplitude stability in the final signal measurement filters (narrow bandwidth extends the loop threshold at the expense of am-

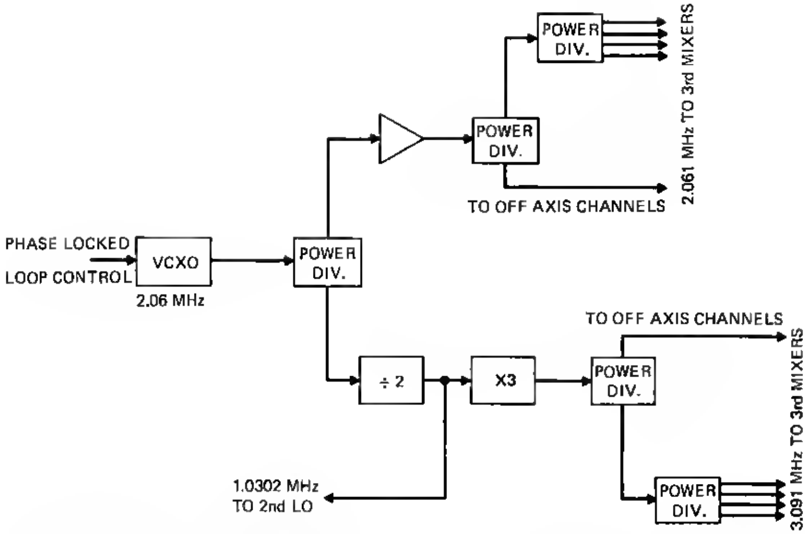


Fig. 8—Receiver third local oscillator in control room [voltage controlled crystal oscillator (VCXO) frequency is 2.0604 MHz].

plitude fluctuations due to FM-AM conversion in the measurement filters). Requirements (*iii*) and (*iv*) guarantee good frequency memory during signal fades and recovery from them without long loop pull-in times.

The loop components are shown in more detail in Fig. 9. The received 19-GHz V copolarized signal, at 325 Hz, is filtered by an active bandpass filter, limited, and phase compared with stable 325-Hz reference. The phase detector output is filtered by an active lowpass filter and used to control the third LO VCXO frequency. Thus, the 19-GHz V-copolarized signal is phase locked to the 325-Hz reference, and with it all other receiver channels.

The PLL noise bandwidth is adjustable over the range 5 to 50 Hz to encompass the range of short-term satellite oscillator stabilities expected. A loop damping factor ζ of 0.8–1 was desired to avoid phase overshoot.[9] Both of these parameters are influenced by both the loop filter and the IF bandpass filter. A test using prototype RF and PLL hardware showed that an overall loop bandwidth B could be achieved with a loop filter natural frequency ω_n (rad/sec) = 1.2 B, a loop damping factor $\zeta = 1.2$, and an IF filter bandwidth = 1.5 B. To achieve the desired 5–50 Hz overall bandwidth range, the loop filter ω_n may be varied (through adjustment of R_1 and R_2 in Fig. 9) over 6–60 rad/sec holding $\zeta = 1.2$, and the IF filter bandwidth varied over 9–90 Hz, holding its gain constant. This adjustment is made in 1 dB switched steps (bandwidth ratio = 1.26).

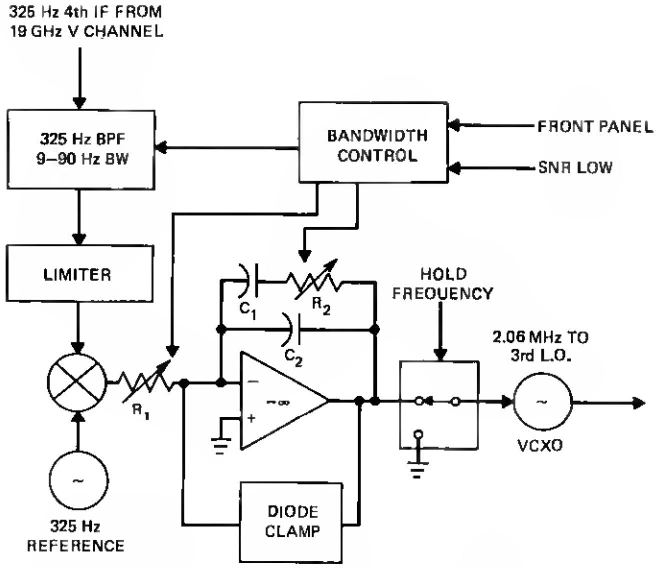


Fig. 9—Receiver third local oscillator PLL circuitry.

The capacitor C_2 in the loop filter rolls the loop off to prevent the sum-frequency (650 Hz) phase detector output from modulating the VCXO. The diode clamps limit the frequency excursion to ±50 Hz. The VCXO control line is grounded on external command, and holds the VCXO at its resting frequency during signal outages.

5.5 Fourth LO

The fourth LO shown in Fig. 10 is self-contained; it generates sideband receiver-channel local oscillators, LOUs and LOLs, which include the required fourth LO frequency corrections. Because of the low frequencies involved and the ease of frequency division using medium scale integrated circuits (ICs) all fourth LO frequencies are derived by division from a single 10.5333-MHz crystal oscillator. The divisions ($\times\, n/128$) that determine the final output frequency ratios are done by digital rate multipliers. These rate multipliers produce n output pulses for every 128 input pulses with n selected by pin connections on the ICs. The n output pulses are not evenly spaced because they are produced by gating the input pulse train. The uneven spacing is equivalent to a deterministic timing jitter on the output waveform which is a fixed multiple of the input pulse period. The three decade dividers (÷ 1000) following the rate multipliers preserve this timing jitter, which is then a much smaller fraction (1/1000) of the output waveform period. In effect this ÷ 1000 reduces the index of the jitter phase-modulation by 1000. The jitter

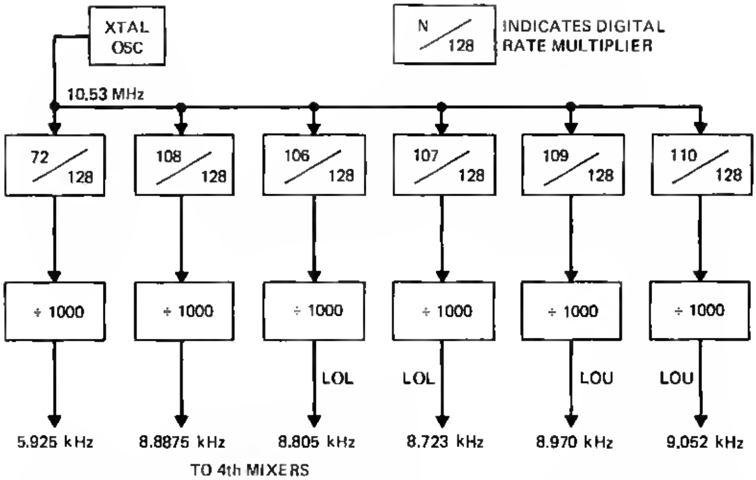


Fig. 10—Receiver fourth local oscillator in control room (crystal oscillator frequency is 10.533333 MHz).

sidebands on the worst fourth LO output (the 9.052 kHz) are 60 dB below the LO signal level; most LO outputs have sidebands >65 or 70 dB below the signal level.

5.6 Frequency acquisition

Since this receiver is designed for unattended operation, it must be capable of automatic frequency reacquisition after signal outages. This is accomplished by sweeping the receiver through the expected range of signal frequencies whenever tracking cannot be maintained. When the signal is located, this sweep stops and the receiver returns to its normal tracking mode.

The receiver's acquisition speed is increased by making use of the signal's last known frequency, its maximum drift rate, and its maximum daily frequency excursion. Let:

$$f_o = \text{last known frequency at time } t_o$$

$$\Delta = |\text{maximum drift rate}|$$

$$f_{\min} = \text{minimum observed frequency}$$

$$f_{\max} = \text{maximum observed frequency.}$$

Then the frequency region to be searched is bounded by

$$\max\{f_{\min}, (f_o - \Delta(t - t_o))\} \le f(t) \le \min\{f_{\max}, (f_o - \Delta(t - t_o))\}.$$

Acquisition speed is also improved for short-duration signal outages by holding the last known signal frequency and inhibiting the frequency

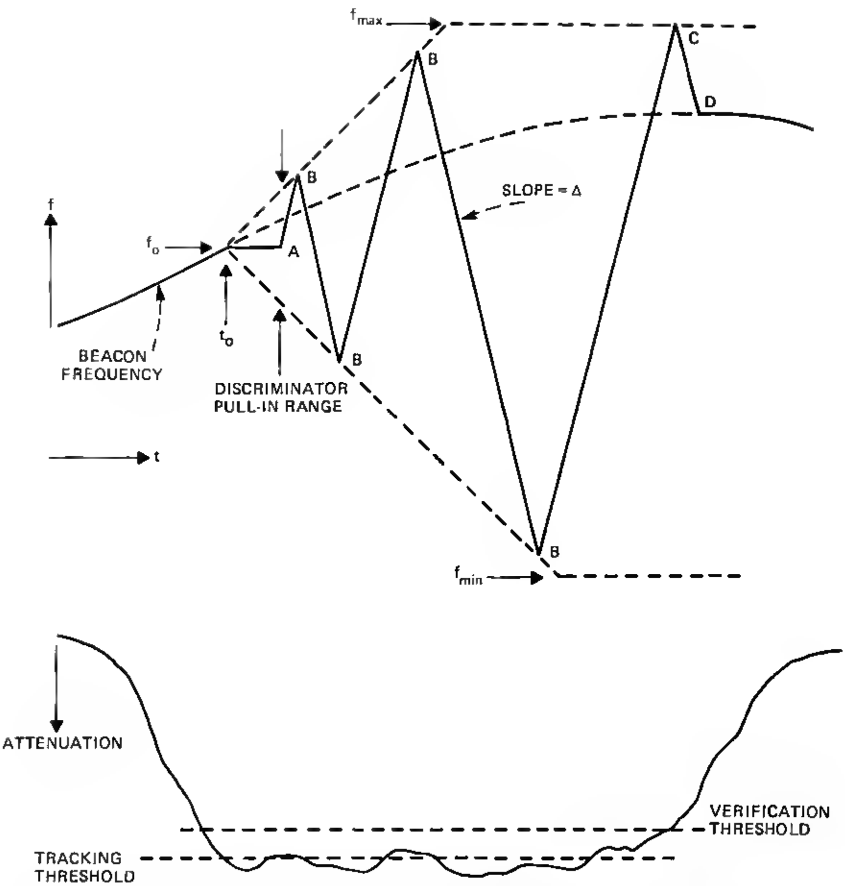


Fig. 11—Typical acquisition sequence for receiver second local oscillator. Frequency tracking is lost at time t_o. Reacquisition frequency search begins at point A: signal is reacquired at point D.

sweep until the maximum frequency uncertainty exceeds the pull-in range of the discriminator.

Figure 11 illustrates a typical receiver acquisition sequence. At time t_o the received signal level fades below the tracking threshold. The register controlling receiver frequency is held at its last value f_o and a counter is counted at a rate Δ to indicate the maximum frequency uncertainty as time progresses. When this uncertainty exceeds the pull-in range of the tracking discriminator (point A) the frequency-control register is stepped up or down continuously to sweep the receiver frequency. Whenever this register's excursion from f_o equals the maximum frequency uncertainty the stepping direction of the frequency control register is reversed, reversing the direction of frequency sweep (points

B). Sweep direction is also reversed at f_{max} and f_{min}, the daily frequency excursion limits (point C). These limits are set by thumbwheel switches, and are reset periodically to account for long-term aging of beacon and receiver oscillators.

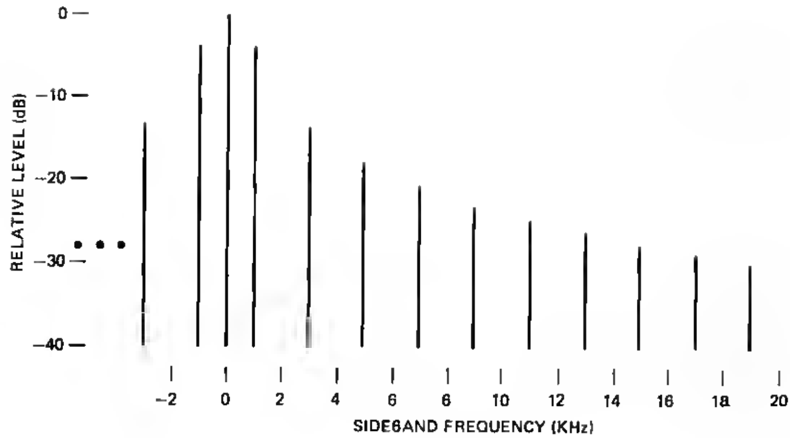
During this frequency sweep the unswitched 19-GHz vertically-polarized channel is observed to detect the presence of the beacon signal. When its presence has been verified (point D) the receiver returns to its tracking mode. The procedure used to accomplish this verification is described below.

As mentioned previously, the 19-GHz received signal is used for reacquisition. However, the situation is complicated by the 1-kHz polarization switching. This switching produces sidebands separated from the carrier by multiples of the switching frequency. The receiver uses knowledge of the relative levels of the carrier and sidebands to discriminate between tbe two.

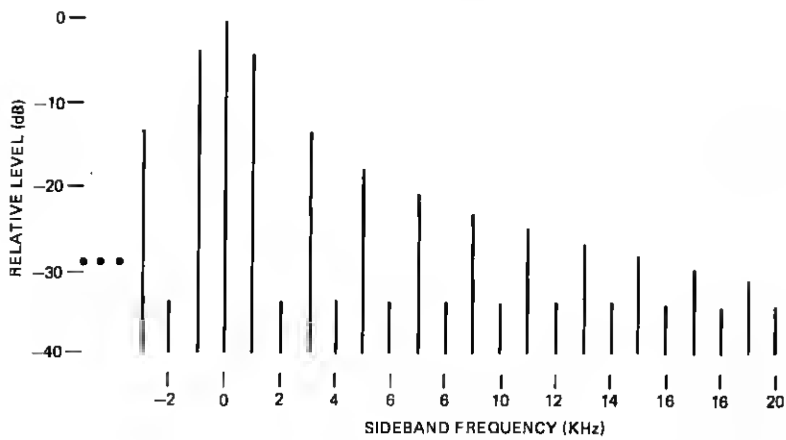
If the polarization modulation were symmetrical (50 percent duty cycle) the received spectrum would appear as in Fig. 12a. Only odd-order sidebands are present; the first-order (±1 kHz) sidebands are down 3.9 dB. This pattern of three signals could be detected using three narrowband filters with 1 kHz spacing to indicate the presence of carrier in the center filter. If modulation asymmetry is considered, however, these three filters are not sufficient. Figure 12b shows the received spectrum for 2 percent modulation asymmetry. This asymmetry has generated even-order sidebands about 4 dB below the level of the 17th and 19th sidebands; three-filter detection would indicate carrier acquisition at these two points. A fourth filter located 2 kHz from the carrier filter, however, resolves this ambiguity. Its relative output will be low for a true detection and high otherwise. Since the asymmetry of the beacon modulation was not known during receiver construction, four filters were used to unambiguously detect carrier acquisition.

This four-filter acquisition detector is shown in Fig. 13. Four narrowband active filters are driven by the 6.25-kHz signal (derived from the 19-GHz V channel) used by the AFC loop. These filters have a two-pole response with matched 30-Hz bandwidths, and are tuned to pass the carrier at 6.25 kHz, two ±1 kHz sidebands, and the −2 kHz sideband. These filters also are contained in a temperature-controlled oven. Since detection decisions are based on ratios of these signals, the filter outputs are envelope detected and logged. Differences of these logged signals then indicate the desired ratios.

The 30-Hz filter bandwidth was chosen as a compromise between sweep speed and SNR at the filter outputs. The filter outputs will not reach their full levels if the sweep speed exceeds an appreciable fraction of filter bandwidth per filter impulse response time. Thus, the maximum permissible sweep speed increases with the square of the detection filter



(a)



(b)

Fig. 12—19-GHz polarization-switching sideband levels, with and without switching asymmetry. Note the even-order sidebands present with switching asymmetry.

bandwidth. The signal level required to achieve a given SNR, however, increases linearly with filter bandwidth. A 10-Hz filter bandwidth (that used in the receiver's 19-GHz signal channels) was found to require a sweep speed of <50 Hz/sec for reliable detection. With the 30-Hz filters used, the sweep speed may be increased to *250* Hz/sec, speeding the reacquisition process.

Several conditions must be satisfied to indicate the presence of received carrier in the 6.25 kHz carrier filter. The carrier filter output must be >10 dB above its no-signal level to assure a reasonable false-alarm rate. Both ±1 kHz sideband filter outputs must be between 2 and 6 dB

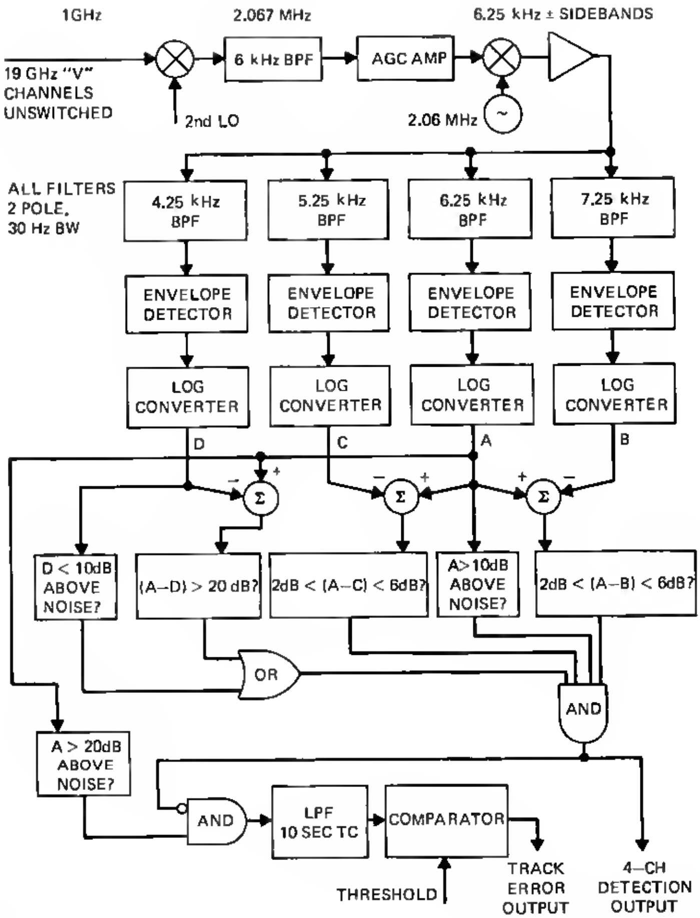


Fig. 13—Four-filter acquisition detector (part of receiver second local oscillator AFC circuitry).

down from the carrier level. This range allows for slight nonlinearities in the log converter slopes and the effect of noise at low signal levels. Finally, the −2 kHz sideband filter output must be either <10 dB above its noise level or >20 dB below the carrier filter output. If all these conditions are satisfied a four-channel detection signal is given. This action interrupts the acquisition frequency sweep and turns on the AFC to attempt to track the signal. The four-channel detection signal is observed for 15 seconds to verify the detection. If the detector output is true for >50 percent of this time, a valid signal is assumed to be present and the receiver returns to its tracking mode. Otherwise, the acquisition search is continued until the signal is found.

A second signal is generated in conjunction with the four-channel

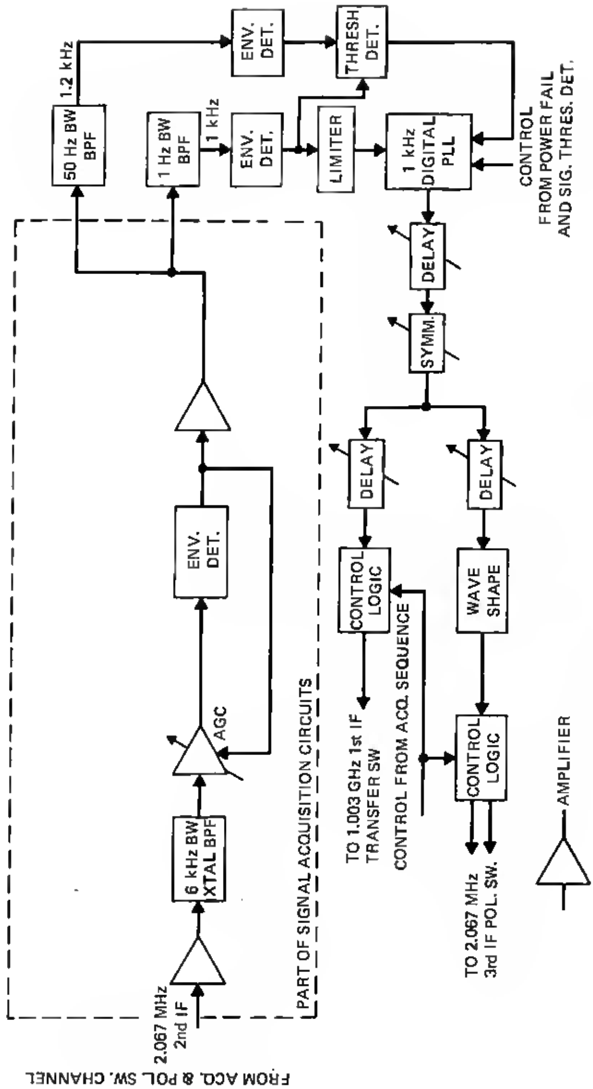


Fig. 14—1-kHz polarization switch synchronizing circuits.

detection output to further guard against tracking of a 19-GHz sideband. An error signal is generated if the carrier filter output is >20 dB above its noise level but no four-channel detection has been made. (Such a condition would occur if the receiver were tracking a sideband.) If this error signal persists for 10 seconds a frequency sweep is initiated over the full uncertainty band. This action should never be performed, but further assures the eventual acquisition of only the 19 GHz carrier.*

VI. POLARIZATION SWITCH SYNCHRONIZATION

A separate IF channel without switches is provided for synchronizing the main-channel polarization switches (see Fig. 1) with the 1-kHz polarization-switched beacon signals. Since the clear-air signal-to-noise ratio in a 2-kHz bandwidth is <+30 dB and low jitter (<5 percent of a switching period) switching is desired for rain attenuation of at least 40 dB, a very long (~100 sec) time constant (narrowband) PLL is required to recover the 1-kHz switching signal. Such a long time constant loop has an even longer pull-in time; therefore, frequency and phase memory through severe rain attenuation events (>40 or 50 dB) on the order of 10 minutes is needed to prevent loss of data for long periods after such severe events. The long PLL time constants and even longer memory are easiest to implement digitally.

The polarization switch synchronizing circuits in Fig. 14 follow the 1 GHz to 2 MHz frequency conversion in the acquisition and polarization switch synchronization channel in Fig. 1. The 1-kHz digital PLL is shown in more detail in Fig. 15.

As indicated in Fig. 14, the automatic gain controlled (ACC) 2-MHz IF amplifier holds the 1-kHz signal into the oven-enclosed 1-Hz BW active BPF constant over the range from the clear air signal level to the noise level of the 1-Hz BW BPF input. The filtered 1 kHz is limited, fed to the PLL, and is detected to provide a threshold comparison voltage. The noise voltage in a 50-Hz BW at 1.2 kHz is rectified and lowpass-filtered for the threshold detector reference.

The square-wave PLL output is phase-locked to the satellite polarization switch. However, because of filter delays, etc., the square-wave transitions and the beacon signal transitions at the receiver switches do not occur at the same instants of time. Delay and symmetry adjustments after the PLL permit adjustment of the switching waveform transitions for coincidence with the signal transitions. Further shaping of the switch driving waveform for the 2-MHz third IF polarization switches (Fig. 1) provides a "dead zone" around transitions to allow switching transients to settle.

* This circuitry was later disabled since it was often triggered by strong depolarization produced by atmospheric ice crystals.[12]

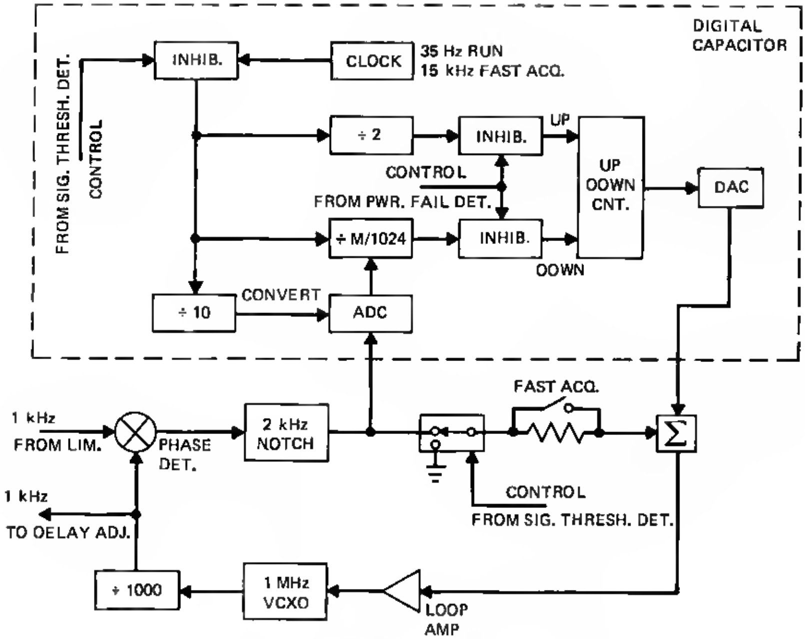


Fig. 15—1-kHz digital phase-locked loop (PLL).

6.1 Digital phase-locked loop

The digital PLL in Fig. 15 is equivalent to a conventional second order loop[9] with an active loop filter. The frequency of a 1-MHz voltage controlled crystal oscillator (VCXO) is controlled by the loop amplifier output voltage. The 1 MHz is divided in digital decade dividers to 1 kHz and compared in phase with the limited 1 kHz derived from the polarization-switched beacon signal. The notch filter following the analog-multiplier phase-detector prevents the 2-kHz sum frequency from saturating the loop amplifier. The straight-through path from the notch filter to the summer (through the switch and resistor described later) establishes the high-frequency open-loop gain. The digital capacitor output is summed with the straight-through path to determine the open-loop low-frequency response.

Operation of the digital capacitor centers on the up-down counter. A fixed clock frequency is divided by two and applied to the count up input of the counter. Input pulses are blocked by the inhibit gates when a control voltage appears on the inhibit control line; otherwise, all input pulses are passed on to the output. The analog-to-digital converter (ADC) samples the phase detector output every 10 clock cycles and converts it to a 10-bit digital word, M, that serves as the control word for a digital rate multiplier. With an ADC input of 0, $M = 512$, and the rate multiplier

passes 512/1024 = 1/2 of the input pulses to the count down input of the counter. Under this condition, the counter alternately counts up and down so there is no net change in count. The pulses into the count up and count down inputs and the convert pulses to the ADC are appropriately timed to prevent ambiguities that would arise if these pulses were nearly coincident. Positive phase detector outputs produce $M > 512$. With $M > 512$ there are more down counts than up counts in a given time period and the counter counts down at a rate proportional to the phase error. With a negative phase error, $M < 512$, there are more up counts than down, and the counter counts up. The count in the counter is analogous to charge stored in a capacitor; M is analogous to current into the capacitor; and the ADC is equivalent to a voltage to current converter. The analogy is completed by the digital-to-analog converter that converts the counter count to a voltage, the capacitor voltage, with 0 count being maximum positive voltage, maximum count being maximum negative voltage and mid-count being 0. The equivalent capacitance is a function of the clock rate, the ADC voltage-to-M factor and the DAC count-to-voltage factor. The counter saturates at maximum count and at 0 count so the contents are not "spilled" by either a continuous positive or negative phase error. The DAC output is filtered to smooth the voltage steps.

In the normal run mode of the PLL, the digital capacitor clock frequency is 35 Hz, the natural frequency,[9] ω_n, of the PLL is 6×10^{-3} and the damping is 0.5. A faster loop response is available for faster acquisition (acq) of phase lock with high signal levels. This is accomplished by increasing tbe clock frequency to 15 kHz to decrease "capacitance" and by decreasing the loop summing resistance to increase the high-frequency open-loop gain and thus maintain the damping at 0.5. The fast acquisition mode can be selected manually with a single switch. The digital capacitor departs from a real capacitor when capacitance is changed by changing clock frequency. Changing clock frequency does not instantly change the counter count; it changes only the rate of count. Thus, digital capacitor voltage is conserved with capacitance change and "charge" is not conserved. This is a useful property in this PLL because the VCXO control voltage and thus the VCXO frequency and phase remain continuous when the loop natural frequency is changed discontinuously. The PLL then remains in lock and does not experience a phase step in going from the fast acquisition mode to the run mode.

Phase and frequency memory during loss of signal or loss of primary power is provided by the up-down counter, loop amplifier, VCXO, and $\div 1000$. These components are supplied from a battery power supply tbat normally floats on a charger.

For loss of power all inputs to the up-down counter are inhibited to prevent start-up transients from disrupting the held count.

For loss of signal, counter memory is provided by inhibiting the clock. Continued operation at the average phase and frequency before the hold is insured by grounding the summing input from the phase detector. The VCXO is in an oven to insure that the stability of the held phase is better than the phase stability of the polarization switching oscillator in the satellite.

VII. DATA COLLECTION

The data collection equipment is common to all receiving channels as indicated in Fig. 16. Data that are critical for maintaining continuity in the data base for long term statistics, e.g., signal attenuation and depolarization, are recorded continuously on analog ink-pen paper-chart recorders. These chart recordings provide a backup in the event of failure of the digital recording system and also provide a "quick look" at the recorded data. The logarithms of signal amplitudes are recorded on the chart recorders with a range of 50 dB.

All receiver outputs are multiplexed along with (*i*) system status indicators such as whether the frequency control loop is tracking or holding in a signal fade, (*ii*) outputs from weather instruments such as rain gauges, thermometers and wind speed recorders, (*iii*) outputs from the on-going interim experiment[10] using the COMSTAR beacon at 128°W, and (*iv*) another propagation experiment[11] using a 12-GHz beacon on the NASA/Canadian Communications Technology Satellite (CTS). These multiplexed signals are digitized, temporarily stored in the minicomputer core memory, screened for relevance, and stored on digital magnetic tape. Multiplexer and analog-to-digital converter sequencing, digital data screening and buffering, and digital tape drive control are handled by the same minicomputer that points the receiving antenna.

The objectives of the data screening procedure are to minimize the amount of superfluous data stored while not losing any relevant propagation data. The screening algorithm copes with the multiplicative signal fluctuations caused by the atmosphere and with the additive noise that dominates at low signal level.

The data taking and screening procedure is outlined in the simplified flow chart in Fig. 17. Four times a second all receiver outputs are digitized and temporarily stored as a sample set (boxes 2 and 3). Each entry j in the sample set is checked for a large impulse change by testing for

$$|\text{sample value}_j - \text{old mean value}_j| < k_{ij}\sigma_{sj}$$

where the old mean value is the stored mean from the preceding time interval, k_{ij} is a scaling factor and σ_{sj} is the expected standard deviation for an individual sample in entry j. For all amplitude entries

$$\sigma_{sj}|_{\text{amplitude}} = \sigma_{aj} + (\text{old mean value}_j) \cdot C_{mj}$$

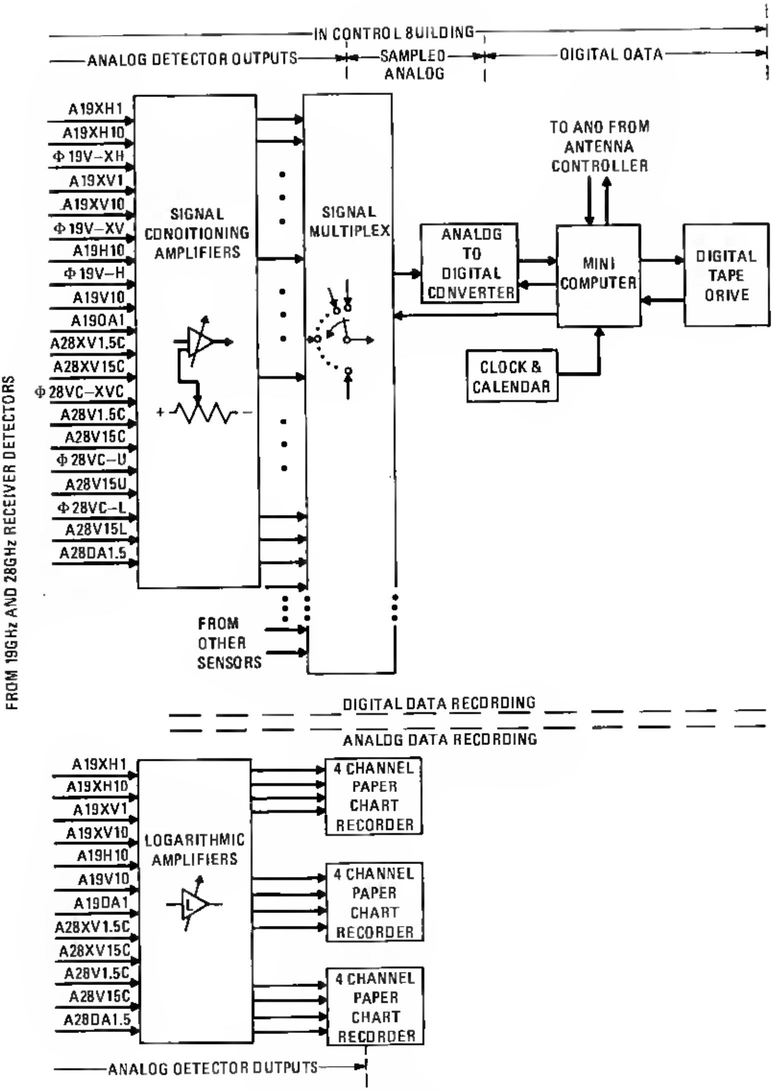


Fig. 16—Data collection equipment.

where σ_{aj} accounts for additive noise and C_{mj} weighted by the old mean value accounts for the multiplicative atmospheric effects. For phase difference entries, the multiplicative factor is inversely scaled by the old amplitude mean of the weakest signal in the phase difference pair:

$$\sigma_{sj}\big|_{\text{phase}} = \sigma_{aj} + C_{mj}/(\text{old amplitude mean}).$$

If an impulse, i.e., a large change in the data, has not occurred, the entries

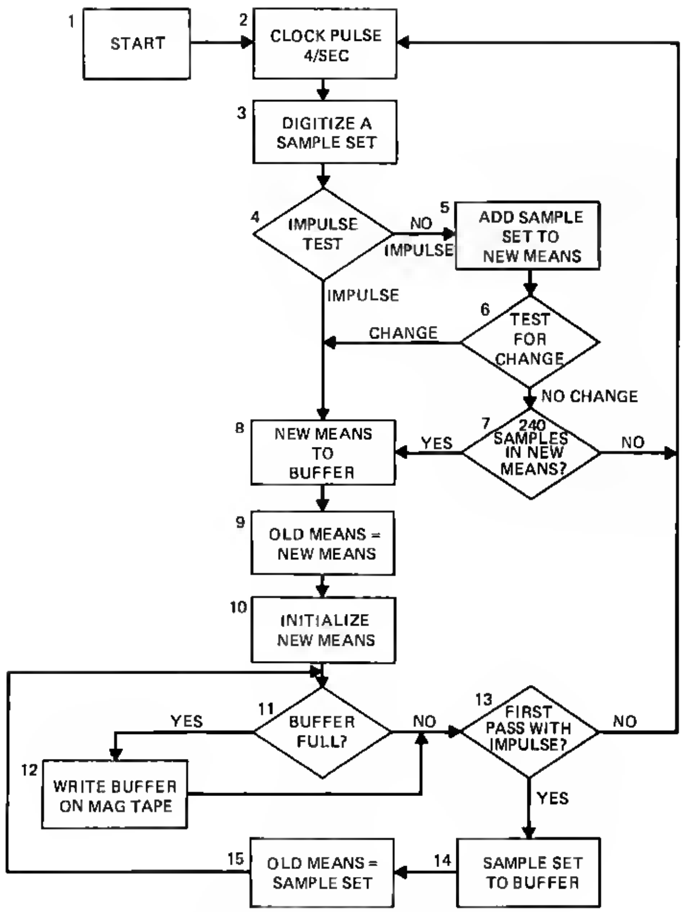


Fig. 17—Flow chart for real-time data screening software.

in the sample set are used to update a set of new mean values being accumulated (boxes 4 and 5). The set of new mean values is then tested for change by

$$|\text{new mean value}_j - \text{old mean value}_j| < k_{mj}\sigma_{sj}/\sqrt{N} \tag{6}$$

where k_{mj} is a different (smaller) scaling constant and N is the current number of samples included in the new mean (box 6). The sample standard deviation σ_{sj} is scaled by $\sqrt{N}$ for this test since the standard deviation of the statistical fluctuations of the mean decreases by this factor (assuming independent samples). This second test quickly detects gradual signal changes. If a change is not detected and if the new mean has been accumulating for less than 1 minute, i.e., $N < 240$, then the data taking program waits for the next clock pulse (box 7). If a change is de-

tected by either of the above tests or if $N = 240$ then the new mean values are transferred to the next unused position in an output buffer (box 8), the old mean values are set equal to the new mean values (box 9) and the new mean values are initialized (box 10). If 20 entries of sets of values have been made into the output buffer, the buffer is full and it is written onto magnetic tape (boxes 11 and 12). If this pass through the procedure is not from the impulse test or is the second pass of the first impulse detected, the program waits for the next clock pulse (box 13). If this is the first pass from a detected impulse, however, the entire set of samples is transferred to the output buffer and the old mean values are set equal to the sample set (boxes 14 and 15). Thus, when an impulse occurs, both the previous running mean and the sample set containing the impulse are stored. Data screening then proceeds normally.

Each set of samples or means recorded also contains the time it was recorded so the time interval spanned is available for off-line data reduction. The data screening software includes provisions for: (*i*) stopping and starting data collection, (*ii*) recovering from primary power interruption, and (*iii*) recording calibration signals on the data tapes.

VIII. CALIBRATION SOURCE

The operation of the entire receiving system may be verified using a 19- and 28-GHz signal source with RF characteristic closely duplicating those of the beacon itself. Knowledge of the signal source amplitude also allows determination of the absolute received signal level. This source was extremely useful during prelaunch receiver construction and check out.

A block diagram of the calibration source (beacon simulator) is shown in Fig. 18. A 66.11-MHz crystal oscillator drives a multiplier chain, producing three output frequencies: 264.44 MHz, 19.04 GHz, and 28.56 GHz. Samples of the 19- and 28-GHz signals are made available to indicate the power output of the source. The 28-GHz signal is attenuated and phase-modulated by either the 264-MHz multiplier output or that output doubled to 529 MHz. The modulation index is set to produce sideband levels similar to those generated by the satellite. A precision calibrated variable attenuator allows level control from near clear-air levels to the receiver noise floor. A cross-polarized signal component is obtained from a 20-dB directional coupler.

The 19-GHz multiplier output is attenuated through fixed and variable attenuators. A cross-polarized component is generated with a 30-dB coupler. Both direct and cross-polarized signals are power-split using 3-dB quadrature hybrids. The two 19-GHz outputs are switched alternately between a "direct" and a "cross-polarized" hybrid output using PIN diode switches. A 1-MHz crystal oscillator and ÷1000 frequency

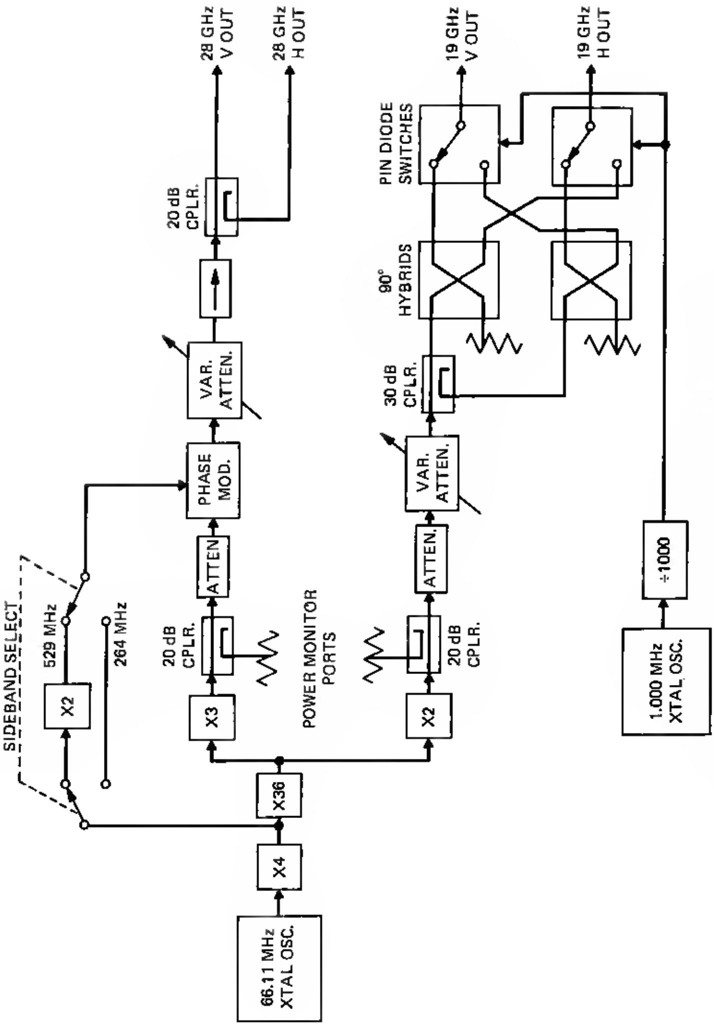


Fig. 18—Receiver calibration source (beacon simulator).

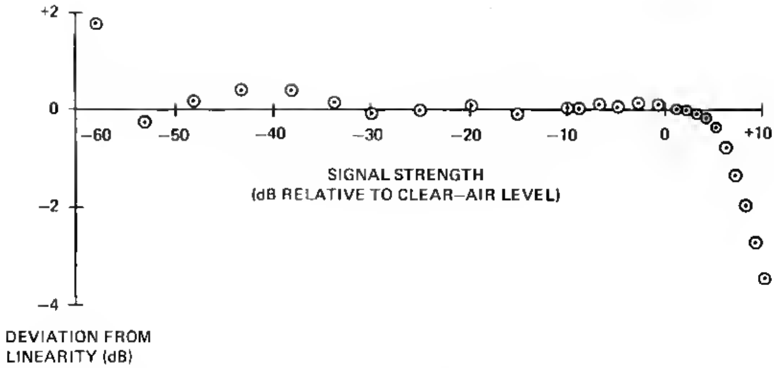


Fig. 19—Receiver input-output amplitude linearity (19-GHz vertical copolarization channel).

divider switch these switches at a 1-KHz rate. Direct and "cross-polarized" signals then appear alternately at the two 19-GHz outputs.

All four source outputs are permanently connected to their respective receiver mixer inputs through directional couplers so that the source signals can be injected with no waveguide switching. The source outputs are disabled by turning off power amplifiers in the frequency multiplier.

Since the calibration source operates in close proximity to the antenna feed frame, care was taken to hold any RF leakage from the source to <-150 dBm. Coupled with the directivity of the antenna feed assembly, this assures precise calibration signal levels down to the system noise floor.

IX. RECEIVER PERFORMANCE

The entire receiver has met all original performance requirements. Care in the design of individual assemblies has allowed integration of the entire receiving system with extremely minimal interaction or other problems. The receiver has been extremely reliable; during more tban one year of operation, no significant data has been lost due to receiver failures. More detailed observations of actual receiver performance are given below.

Overall receiver linearity is nearly perfect. Figure 19 shows the deviation from perfect linearity of a typical receiver output (19-GHz vertical copol) as a function of input signal level. Other receiver outputs exhibit similar results. The calibration source was used to obtain these results; for the upper 20 dB of input level the source was temporarily connected directly to the mixer inputs. The input signal level is normalized to the clear-air satellite input level. The receiver is seen to be linear within ±0.4 dB for all expected input levels down to 55 dB at-

tenuation. Below this level the measurement becomes noise-dominated. The 1-dB gain compression point occurs 6.5 dB above the clear-air input level. The receiver exhibits negligible differential phase shift with variations of signal amplitude. As an example, the 19-GHz copolarized differential phase is constant within 0.1° from 4 dB above the clear-air input level to the 40-dB attenuation level. Two degrees of phase error is observed at the 50-dB attenuation level.

The receiver has shown excellent long-term gain stability. Measurements over a 4-month period have shown a receiver output change of less than 0.2 dB. This figure includes both receiver and antenna gain changes and satellite power fluctuations.

Long-term measurements of 19 GHz copolarized differential phase are stable to within ±2°. This indicates the overall phase stability of both the satellite and receiver antennas and much of the receiving electronics, including all narrowband filters. Early measurements on the antenna feed frame[4] showed a temperature coefficient of <0.2°/C° at 19 GHz. This has since been improved through equalization of waveguide lengths.

The 2-MHz crystal filters have a temperature coefficient of <0.16°/C°. The two critical 19-GHz V and H copolarized filters were chosen for matched temperature coefficients, and track within ±0.2° over 15 to 40°C. The air-conditioned building environment maintains phase tracking on all other channels within ±1°.

The third and fourth IF active bandpass filters are in temperature-controlled ovens and are unaffected by ambient temperature variations. These filters were bandwidth-matched to within ±1 percent, yielding a frequency-induced phase error of <0.5° for static frequency offsets of up to 5 Hz.

As stated previously, the receiver noise floor is set by front end noise; all following amplifiers and frequency conversions contribute <0.1 dB to the overall system noise level.

The polarization switch synchronization circuitry must accurately track the 1-kHz polarization switching phase to prevent leakage of copolarized signals into receiver crosspolarized channels. This circuitry exhibits ±5° phase jitter during a 40-dB fade and holds phase open-loop for greater attenuation. A dead band of ±9° is sufficient to maintain open-loop synchronization for at least 10 minutes.

The AFC loop will reliably track the satellite frequency excursions during a 51 dB fade at 19 GHz (copol SNR ≥ 9 dB). Below this level, the loop initiates its reacquisition sequence. Initial four-channel detection is accomplished with 50 percent probability at the 49 dB fade level, while 50 percent final verification probability is reached at the 47 dB fade level.

An example of data obtained during a typical rain event is shown in

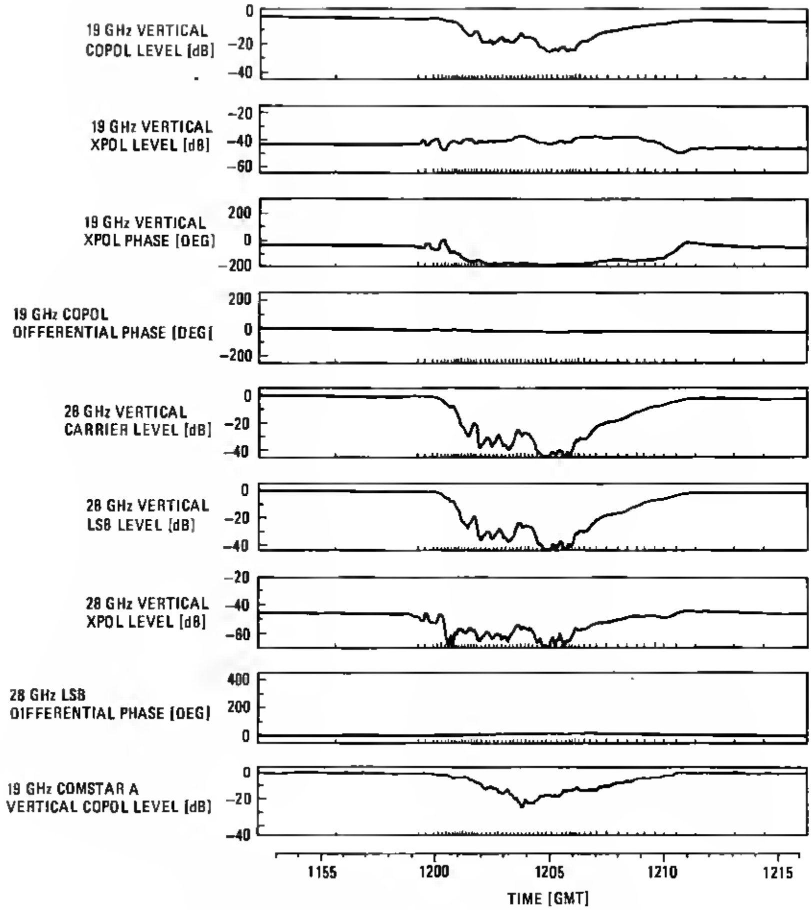


Fig. 20—Typical receiver outputs from data screening software.

Fig. 20. Only 9 of the 25 data channels recorded are shown here. The display was generated by software used for off-line data screening. The ticks along the x-axis delimit records containing 20 sets of data points (see Section VII on data collection) and indicate the increase in data-recording rate during such events.

The 4 upper traces display 19-GHz vertical copolarized (A19V) and cross-polarized (19XV) signal levels, vertical cross-polarization phase (ϕ19V-XV), and copolarization differential phase (ϕ19V-H). The cross-polarization phase zero reference was obtained by rotating the antenna feed frame slightly to leak a copolarized component into the cross-polarization channels.

The next four traces display 28-GHz vertical copolarized carrier (A28VC), lower sideband (A28VL), and cross-polarized signal level

(A28XV), and lower sideband differential phase (ϕ28VC-L). The close agreement of carrier and sideband amplitudes and lack of relative phase shift indicate little medium dispersion over the 264-MHz spacing. The phase fluctuations observed at low signal amplitudes are due to the low measurement SNR, and are not medium effects.

The final trace displays the vertical copolarized output of the colocated interim receiver[10] observing the COMSTAR beacon at 128°W longitude. The differences between this and the first trace result from the different ray paths taken through the rain.

X. SUMMARY

This paper has presented the design requirements for the Bell Laboratories Crawford Hill 19- and 28-GHz COMSTAR beacon receiving electronics, and has described the hardware and techniques used to achieve these requirements. The receiver operates unattended, continuously collecting data to determine attenuation, depolarization, differential phase shift, dispersion and angular scatter produced by precipitation. The coherence of the various transmitted beacon signals allows use of receiver noise bandwidths as narrow as 1.6 Hz to determine depolarization during severe fading. The receiving system meets all the design requirements, and has already collected data reliably for over one year.

XI. ACKNOWLEDGMENTS

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